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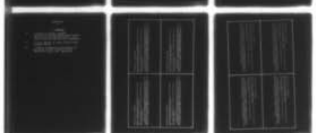

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APPLICATION OF DIGITAL FILTERING TO AN INFRARED INTRUSION ALARM DEVICE

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(6) APPLICATION OF DIGITAL FILTERING

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RESUME

On a étudié théoriquement l'application du traitement de signaux numériques à un système d'alarme d'intrusion et comparé les performances de quatre types différents de filtres-comparateurs. Des simulations en ordinateur nous ont permis d'évaluer la probabilité de détection et le taux de fausses alarmes. (NC)

ABSTRACT

The application of digital signal processing to an intrusion alarm device is studied theoretically. The performance of four different types of filter-comparator units is evaluated and compared. Computer simulations are also carried out to obtain the probability of target detection and the false alarm rate. (U)

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1.0 INTRODUCTION

This report presents the results of a theoretical investigation of the application of digital signal processing to an intrusion alarm device. The investigation was carried out in parallel with the construction of a working model of the intrusion alarm at Laval University under a DRB contract [1]. The theoretical studies and the experimental work involved in the construction of the alarm were both guided by the findings of an earlier DREV project. A description of the main results of this project can be found in a recent patent application by Giroux [2].

Although the prototype of the intrusion alarm operates in the infrared spectral region, the concept described in Ref. 2 is of a general nature. There, the basic principle of the apparatus was very clearly stated: "In accordance with the present invention, apparatus for detection of variations in the radiation pattern from a defined space is provided which comprises means for deriving a digital signal from the scanned scene and means for storing the digital signal as a series of discrete digital values in a digital memory. A digital comparator is provided for comparing the digital output for one scan with the stored digital values in the memory for a preceding scan of the scene. *If there is a discrepancy* exceeding some pre-determined threshold value (to account for noise signals, slow changes in ambient light etc) between the digital values for a given portion of the scanned scene and the stored digital values for the same portion of the scanned scene obtained on a previous scan, then an alarm is actuated".

In this report, we examine theoretically the application of the above principle to the infrared intrusion alarm (IRIA). To do so the first task is to formulate the problem taking into account the practical requirements and constraints. The function of the digital filter and comparator is discussed in Sec. 2.0, after a brief description of the general layout of the IRIA. Details concerning the mechanical and optical components of the device are given in Ref. 1. In Sec. 3.0 the physical quantities based on which various possible filter-comparator units are evaluated are studied. These include the time of response, the system transfer function, the probability of target detection and the rate of false alarm. Comparisons between different filter-comparator units are made. To verify the analytical results of Sec. 3.0 filter-comparator units have been simulated on a computer. The results of the simulation, the performance characteristics and the relative merits of the units considered are described in Sec. 4.0. Section 5.0 summarizes the main results and makes specific recommendations for the implementation of the digital filter-comparator unit to be incorporated to the IRIA. This work was performed at DREV between April 1974 and April 1975 under PCN 33A07 (formerly PCN 15B25, Project 97-82-06) "Intrusion Detection and Alarm."

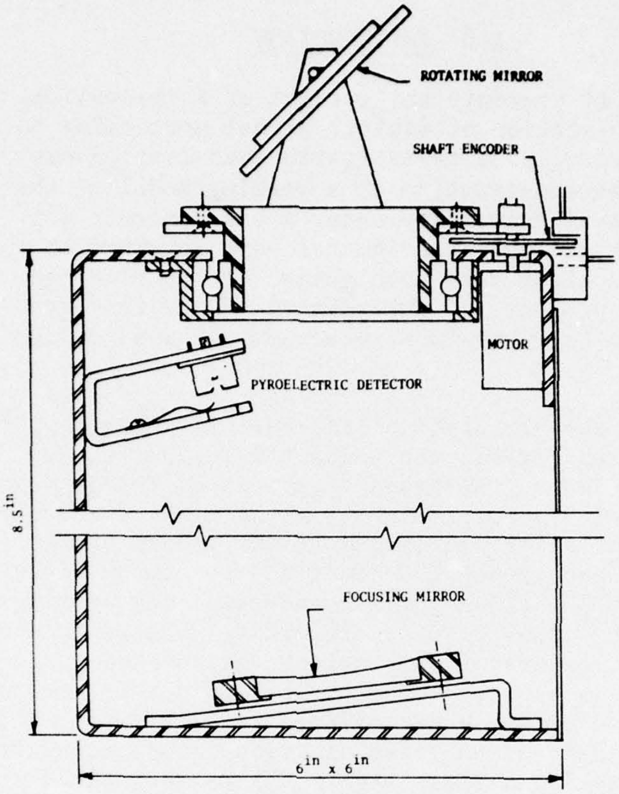

FIGURE 1 - Optics and detector stage

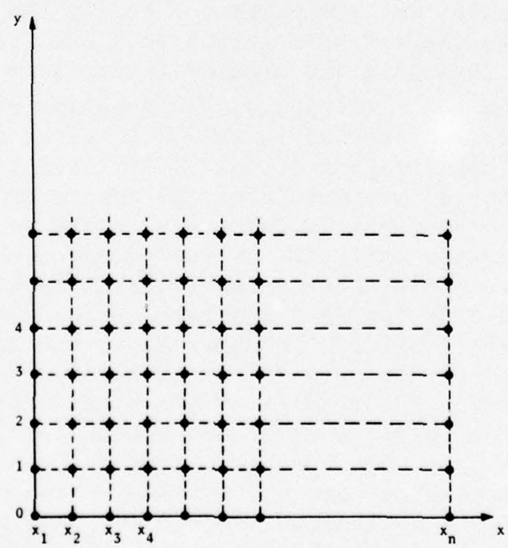

FIGURE 2 - Two dimensional lattice on which M(x, y) is defined

2.0 FORMULATION OF THE PROBLEM

To define the problem of signal processing associated with the IRIA, it is useful to give a brief description of the device. In the IRIA, the incoming signal is obtained by scanning a scene with a plane mirror (Fig. 1). The plane mirror, driven by a motor, can rotate at a maximum angular speed of 6.3 rad/s. The scene is sampled at a rate of 1000 points per revolution. A pyroelectric TGS infrared detector measures the infrared radiation between 2 and 20 μm reflected by the plane mirror and then focussed onto it by a concave spherical mirror. The instantaneous field of view of the instrument is approximately 13.2 mrad by 13.2 mrad. The signal voltage generated at the detector is converted to digital form for each sampling position of the scanning mirror.

For a given scan y ($y \geq 0$), a sequence of digital signals $M(x_1, y)$, $M(x_2, y)$, ... $M(x_n, y)$ is obtained. The quantity $M(x, y)$ is the signal voltage generated when the intrusion alarm scans the spatial element with an angular coordinate x measured relative to an arbitrarily chosen axis perpendicular to the axis of rotation of the scanning mirror. The value of y which enumerates the scans is incremented by unity as the IRIA completes one scan. The signal sequence $M(x, y)$ passes through a digital filter-comparator unit (DFCU) after an initial period of time necessary for stabilizing and establishing the background signal level. The DFCU decides on a continuous basis whether, at any given instant, the incoming signal $M(x, y)$ indicates a sudden significant change in the infrared radiation level incident on the IRIA from the direction x. To carry out this task, information on the background radiation and the noise associated with various sources of fluctuations must be collected, processed, stored and updated as the scanning proceeds. In other words, the DFCU must be designed in such a way that, at any point (x,y) in the two dimensional lattice shown in Fig. 2, it can decide whether $M(x,y)$ signals a significant change in the infrared radiation which may be attributed to an intrusion based on the information obtained before the arrival of $M(x,y)$.

From a practical point of view, it is convenient to divide the signal sequence for all angular positions up to a given scan into sub-sequences. Each sub-sequence corresponds to a given direction x so that the sub-sequences for different directions are processed in parallel independent of one another. In situations where the background scene being scanned can be divided into sectors within each of which the background and the noise characteristics are the same the amount of parallel processing can be reduced. For our discussion we consider the more general situation so that the DFCU can cope with changes in the background in each direction without modification. Consequently, for any given direction x, updated information for intrusion detection must be provided in a well-defined

manner by the DFCU as the latest incoming signal $M(x,y)$ is registered. Furthermore, the required information must be extracted from the signal sequence $M(x,y')$ where $y' = 0,1,2, \ldots\ldots,y-1$.

Consider the case when the IRIA is scanning a stationary scene, i.e., the scene radiance is constant in time. If the effect of noise can be neglected, the digitized signal $M(x,y)$ is independent of y for any value of x. If no intrusion occurs during the (y-1)th scan the signal $M(x,y-1)$ can be used as a reference level against which $M(x,y)$ may be measured. In a real situation, however, the noise from various sources must be taken into account. One immediate result is that no scene can be regarded as strictly stationary. Besides, only the variations in $M(x,y)$ not attributable to the fluctuations due to noise are of interest.

To formulate the problem in a way suitable for our subsequent discussions, let us assume that we have N sub-sequences of incoming signals:

$$M(x_1,0),\ M(x_1,1),\ M(x_1,2),\ \ldots\ldots$$

$$M(x_2,0),\ M(x_2,1),\ M(x_2,2),\ \ldots\ldots$$

$$\cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot$$

$$\cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot$$

and $$M(x_N,0),\ M(x_N,2),\ \ldots\ldots\ldots\ldots\ldots$$

where x_1, x_2,... and x_N are the directions sampled in each scan. Without intrusion, each signal $M(x,y)$ may be regarded as a sum of two components. These are the background signal and the noise component respectively. Given a sub-sequence of incoming signals $M(x,y')$ and $0 \leq y' \leq y-1$ (x is fixed and y' varies between 0 and y-1), and given that no intrusion occurs in the corresponding period of time (y-1)T (where T is the time required to complete one scan), the problem is to decide whether the signal $M(x,y)$ indicates a significant variation not expected from the behaviour of the signals in the sub-sequence $M(x,y')$; $0 \leq y' \leq y-1$ which precedes $M(x,y)$. To do so, the DFCU should be designed to accomplish two things: (i) it should predict the background signal component of $M(x,y)$, from $M(x,y')$; $0 \leq y' \leq y-1$; and (ii) it should provide some suitable statistical measure of the noise components in the signal sequence so that a significance level can be established. In other words, the problem is to extrapolate a time sequence in the presence of noise and, at the same time, measure the statistical properties of the noise. To determine whether $M(x,y)$ indicates a significant change in the measured infrared radiation level, it is compared with the extrapolated background signal. With the known statistical properties of the noise component, one can find the

probability that a given difference between the incoming signal and the background is caused by random fluctuations.

3.0 DIGITAL FILTERING

The previous section considered the problem of signal processing in the intrusion alarm. We assume for our discussion that the time variation in the background signal is slow so that it can be regarded as a constant over several scanning intervals. With this assumption, the problem of extrapolating the background signal becomes essentially one of filtering out the noise component from the incoming signal sequence.

In the patent application [2] a filter has been suggested for implementation. In this section we study this filter (F1, for brevity), along with some other possible alternative filters, in detail.

3.1 Criteria for filter comparison

The final objective for the design of the DFCU is, of course, the maximization of the probability of detection of intrusion and the minimization of the false alarm rate. At the same time, subject to the practical constraints, the IRIA as a whole should be simple and inexpensive to implement. To compare the performance of different filters it is convenient to use the following requirements as criteria:

(a) The time of response of the digital filter should be short.

(b) The filter should efficiently filter out the noise component in the signal sequence $M(x,y)$

(c) The filtered signal sequence should follow the slow variation (i.e., over time intervals >> T) in the background signals without appreciable distortion.

We will see later that the filter which performs best according to these requirements also gives a higher detection probability and a lower false alarm rate.

3.2 Choice of digital filters

In the choice of filters we have to restrict ourselves to those which are simple to implement and need a minimum of storage memory. Here we consider four different types of filters: F1, F2, F3 and F4. Some of the properties of F1 have been given in Ref. 2; the impulse response function of F1 together with that of the other filters

are shown in Fig. 3. If the impulse response function of a given filter is denoted by w (y) and the input signal sequence is $M(x,y')$, $0 \leq y' \leq y$, the output of the filter at time yT, is

$$\overline{M}(x,y) = \sum_{y'=0}^{y} w(y')\, M(x,y-y')$$

$$= \sum_{y'=0}^{y} w(y-y')\, M(x,y') \qquad (1)$$

As usual we assume that the filter is causal, i.e. the output signal $\overline{M}(x,y)$ depends on the inputs from the initial moment up to the moment yT. To simplify our notations we omit the variable x in $M(x,y)$ and $\overline{M}(x,y)$. All the equations involving these quantities are valid for an arbitrary but fixed value of x. Thus, we replace Eq. 1 by:

$$\overline{M}(y) = \sum_{y'=0}^{y} w(y')\; M(y-y') = \sum_{y'=0}^{y} w(y-y')\, M(y') \qquad (1')$$

The impulse response function completely specifies the properties of the filter. The analytical expressions of the impulse response function of the four types of filters are:

$$\text{F1:} \quad w_1(y) = (1-e^{-\alpha})\, e^{-\alpha y} \quad (\alpha > 0) \qquad (2)$$

$$\text{F2:} \quad w_2(y) = 1/y_0 \qquad \text{for } y < y_0$$

$$= 0 \qquad \text{otherwise} \qquad (3)$$

$$\text{F3:} \quad w_3(y) = \left[\frac{1 - e^{-\alpha}}{1 - e^{-\alpha y_0}}\right] e^{-\alpha y} \quad \text{for } y < y_0$$

$$= 0 \qquad \text{otherwise} \qquad (4)$$

$$\text{F4:} \quad w_4\,(y) = e^{-\alpha y^2} \Big/ \sum_{y'=0}^{\infty} e^{-\alpha y'^2} \quad (\alpha > 0) \qquad (5)$$

Both F1 and F4 have impulse response functions of infinite duration. While the impulse response of F1 decays exponentially, that of F4 is of the gaussian type. In Ref. 2, F1 is characterized by a constant K which is related to α by the simple relation $K = e^{\alpha}$. The impulse response functions of F2 and F3 are of finite duration.

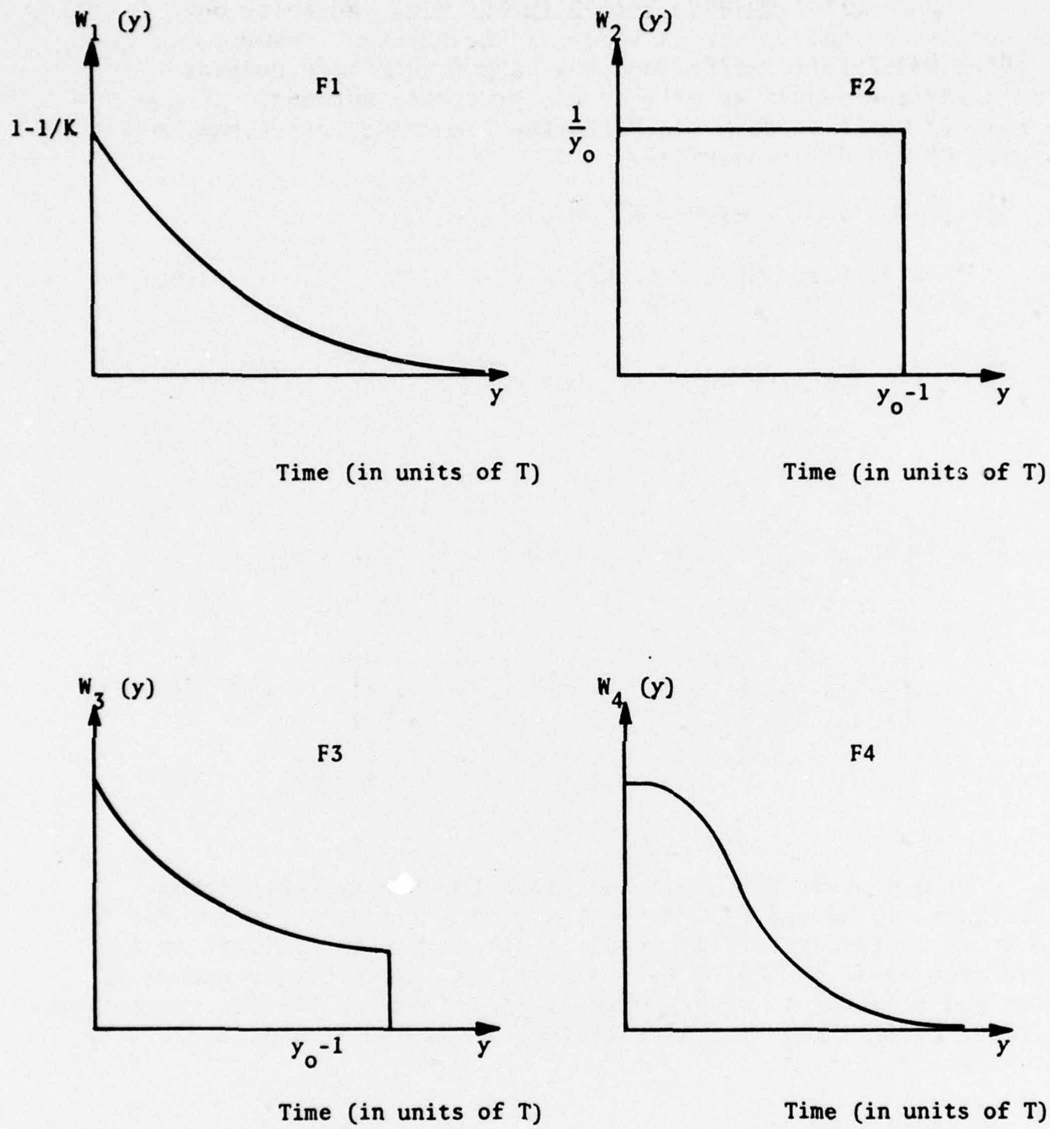

FIGURE 3 - Impulse response functions of filters F1, F2, F3 and F4

Substituting Eqs. 2 to 5 in Eq. 1, we can write down the output of each filter in terms of the inputs. However, it is often possible to write down the output in a more compact form using the inputs as well as the preceding outputs. It can be readily verified that the following recurrence relations hold for F1, F2 and F3 respectively:

$$\text{F1:}\quad \overline{M}(y) = (1 - e^{-\alpha})\, M(y) + e^{-\alpha}\, \overline{M}(y - 1)$$

$$= \left(1 - \frac{1}{K}\right) M(y) + \frac{1}{K}\, \overline{M}(y - 1) \qquad (6)$$

$$\text{F2:}\quad \overline{M}(y) = \frac{1}{y_0}\, M(y) + \overline{M}(y - 1) \quad \text{for } y < y_0$$

$$= \frac{1}{y_0}\, M(y) = \frac{1}{y_0}\, M(y - y_0) + \overline{M}(y - 1) \quad \text{for } y \geq y_0 \qquad (7)$$

$$\text{F3:}\quad \overline{M}(y) = \left[\frac{1 - e^{-\alpha}}{1 - e^{-\alpha y_0}}\right] M(y) + e^{-\alpha}\, \overline{M}(y - 1) \quad \text{for } y < y_0$$

$$= \left[\frac{1 - e^{-\alpha}}{1 - e^{-\alpha y_0}}\right] \left[M(y) - e^{-\alpha y_0}\, M(y - y_0)\right] + e^{-\alpha}\, \overline{M}(y - 1)$$

$$\text{for } y > y_0 \qquad (8)$$

Eqs. 6 to 8 provide the basic equations for the implementation of filters F1, F2 and F3. A similar equation does not exist for F4 and it is necessary to rely on Eq. 1 directly. In contrast to the recurrence equation, Eq. 1 for F4 contains a much larger number of terms and requires a correspondingly larger memory for implementation. In fact, F4 is a non-recursive filter, while F1, F2 and F3 are recursive.

3.3 Response time

The first filter characteristic we wish to consider is the response time. To facilitate the subsequent discussion we shall define the response time τ of a filter as the time required (in units of the scanning time T) for the filter output to rise to within 2% of the value of a time-independent input sequence. The need for such a definition of τ arises because its functional form changes from one filter to another. Since the impulse response functions of the four types of filters we consider are non-negative, the response time is given (in units of the scanning

time T) by the minimum value of y with which the following inequality is satisfied:

$$\sum_{y'=0}^{y} w(y') \geqslant 0.98 \tag{9}$$

For F1, the response time can be determined from the minimum value of y so that:

$$\sum_{y'=0}^{y} (1 - e^{-\alpha}) \quad e^{-\alpha y'} \geqslant 0.98 \tag{10}$$

or

$$e^{-\alpha(y + 1)} \geq 0.02 \tag{11}$$

If τ_1 is the response time of F1 we obtain:

$$\tau_1 = \text{Int} \left[- \frac{\ln \; 0.02}{\alpha} \right] \tag{12}$$

where Int [x] denotes the smallest integer such that Int [x] > x.

From Eq. 3 it is evident, since the impulse response function of F2 is rectangular, that the response time τ_2 is equal to y_0 for $y_0 \leq 50$.

For F3,the time of response τ_3 is given by the least value of y for which the following inequality is satisfied

$$1 - e^{-(y + 1)} \geq 0.98 \quad (1 - e^{-\alpha y_0}) \tag{13}$$

Hence

$$\tau_3 = -1 + \text{Int} \left[\frac{\ln \; (0.98e^{-\alpha y_0} - 0.02)}{\alpha} \right] \quad . \tag{14}$$

For a fixed value of τ_3,more than one combination of the parameters α and y_0 may be found so that Eq. 14 is valid.

The fact that F4 is non-recursive makes it impossible to write down an exact expression for the response time τ_4. By definition,τ_4 is the minimum value of y which satisfies the inequality:

$$\sum_{y'=0}^{y} e^{-\alpha y'^2} \Big/ \sum_{y'=0}^{\infty} e^{-\alpha y'^2} \geq 0.98 \tag{15}$$

To calculate τ_4 approximately, we write:

$$S(m) \equiv \sum_{n=0}^{m} e^{-\alpha n^2} \tag{16}$$

$$S_1(m) \equiv \sum_{n=1}^{m+1} e^{-\alpha n^2} = S(m) - 1 + e^{-\alpha(m+1)^2} \tag{17}$$

and $$S(m) \equiv \int_0^m e^{-\alpha x^2} dx \tag{18}$$

The integral S (m) can be expressed as:

$$S(m) = \frac{1}{\sqrt{2\alpha}} \int_0^{\sqrt{2\alpha m}} e^{-t^2/2} = \sqrt{\frac{\pi}{2}} \left\{ P(\sqrt{2\alpha m}) - 0.5 \right\} \tag{19}$$

where P(x) is the error integral:

$$P(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{x} e^{-t^2/2} dt$$

Using the approximation:

$$S(m) \simeq \frac{1}{2} \left[S(m) + S_1(m) \right] \tag{20}$$

and Eq. 19, we have:

$$S(m) = \sqrt{\frac{\pi}{\alpha}} \left\{ P(\sqrt{2\alpha m}) - 0.5 \right\} + 0.5 \left[1 - e^{-\alpha(m+1)^2} \right] \tag{21}$$

The above equation together with the fact that:

$$S(\infty) = 0.5 \left(1 + \sqrt{\frac{\pi}{\alpha}} \right)$$

enables us to write Eq. 15 as:

$$\sqrt{\frac{\pi}{\alpha}} \left\{ P(\sqrt{2\alpha m}) - 0.5 \right\} + 0.5 \left[1 - e^{-\alpha(m+1)^2} \right]$$

$$\geq 0.98 \times 0.5\left(1 + \sqrt{\frac{\pi}{\alpha}} \right) . \tag{22}$$

The least value of m satisfying Eq. 22 gives the time of response τ_4. Equation 22 gives a sufficiently accurate method for calculating τ_4 without having to calculate the series in Eq. 15 directly by term-wise summation.

3.4 Frequency response and phase transfer functions

Up to this point, our discussion of the various types of digital filters has been restricted to the time domain. However, it is instructive to study the behaviour of the filters in the frequency domain. The mathematical tool used for this purpose is the Z-transform. For any time sequence, $\left\{N(y)\right\}$ the Z-transform is defined by:

$$W_N(z) = \sum_{y=0}^{\infty} N(y) . z^{-y} \tag{23}$$

from which the original sequence can be obtained by the inversion formula:

$$N(y) = \frac{1}{2\pi i} \oint_{\substack{\text{unit} \\ \text{circle}}} W_N(z) \; z^{y-1} dz \quad . \tag{24}$$

If the impulse response function of a causal digital filter is given, the transfer function W(z) of the filter is defined as the Z-transform:

$$W(z) = \sum_{y=0}^{\infty} w(y) \; z^{-y} \tag{25}$$

and the corresponding inversion formula is:

$$w(y) = \frac{1}{2\pi i} \oint_{\text{unit circle}} W(z) \; z^{y-1} dz \tag{26}$$

The properties of a given digital filter are completely specified by its transfer function. Since W(z) is a complex-valued function, one may separate its real and imaginary parts:

$$W(z) = W^R\ (z) + iW^I\ (z) \tag{27}$$

such that functions W^R (z) and W^I (z) are real valued. The frequency response function R(ω) of the filter is given by:

$$R(\omega) = \sqrt{\left[W^R\ (z = e^{i\omega T})\right]^2 + \left[W^I\ (z = e^{i\omega T})\right]^2} \tag{28}$$

where T is the sampling interval. *The phase transfer function* (ω) is given by the formula:

$$\tan\ (\omega) = \frac{W^I(z = e^{i\omega T})}{W^R(z = e^{i\omega T})} \tag{29}$$

From Eqs. 28 and 29 we can see that the frequency response and the phase transfer functions are determined by the behaviour of W(z) on unit circle of the complex z-plane.

For the filters F1, F2 and F3,it is straightforward to calculate the transfer functions as the corresponding series (see Eq. 25) can be summed readily. In fact, for F1:

$$W_1(z) = \sum_{y=0}^{\infty} w_1(y)\ z^{-y}$$

$$= \sum_{y=0}^{\infty} \left(1 - e^{-\alpha}\right) e^{-\alpha y}\ z^{-y}$$

$$= \left(1 - e^{-\alpha}\right) \sum_{y=0}^{\infty} \left(e^{-\alpha}\ z^{-1}\right)^y$$

$$= \frac{1 - e^{-\alpha}}{1 - e^{-\alpha} z^{-1}} \tag{30}$$

For F2:

$$W_2(z) = \sum_{y=0}^{\infty} w_2\ (y)\ z^{-y}$$

$$= \frac{1}{y_0} \sum_{y=0}^{y_0-1} z^{-y} = \frac{(1 - z^{-y_0})}{y_0(1 = z^{-1})} \tag{31}$$

Similarly, one can show for F3 that:

$$W_2(z) = \frac{(1 - e^{-\alpha})\ (1 - e^{-\alpha y_0}\ z^{-y_0})}{(1 - e^{-\alpha y_0})\ (1 - e^{-\alpha} z^{-1})} \tag{32}$$

For F4 we are unable to reduce the transfer function from the definition:

$$W_4(z) = \sum_{y=0}^{\infty} e^{-\alpha y^2}\ z^{-y} \Big/ \sum_{y=0}^{\infty} e^{-\alpha y^2} \tag{33}$$

to a more compact form.

The frequency response functions for F1, F2 and F3 can be obtained from the transfer functions and Eq. 28. They are given by the expressions:

$$\text{F1:} \quad R_1(\omega) = (1 - e^{-\alpha}) \Big/ (1 - 2e^{-\alpha} \cos\omega T + e^{-2\alpha})^{\frac{1}{2}} \tag{34}$$

$$\text{F2:} \quad R_2(\omega) = (1 = \cos y_0\ \omega T)^{\frac{1}{2}} \Big/ y_0\ (1 - \cos \omega T)^{\frac{1}{2}} \tag{35}$$

$$\text{F3:} \quad R_3(\omega) = \frac{(1 - e^{-\alpha})\ (1 - 2e^{-\alpha y_0} \cos y_0\ \omega T + e^{-2\alpha y_0})^{\frac{1}{2}}}{(1 - e^{-\alpha y_0})\ (1 - 2e^{-\alpha} \cos \omega T + e^{-2\alpha})^{\frac{1}{2}}} \tag{36}$$

The frequency response functions for F1, F2, F3 and F4 are compared for given times of response in Figs. 4 to 6.

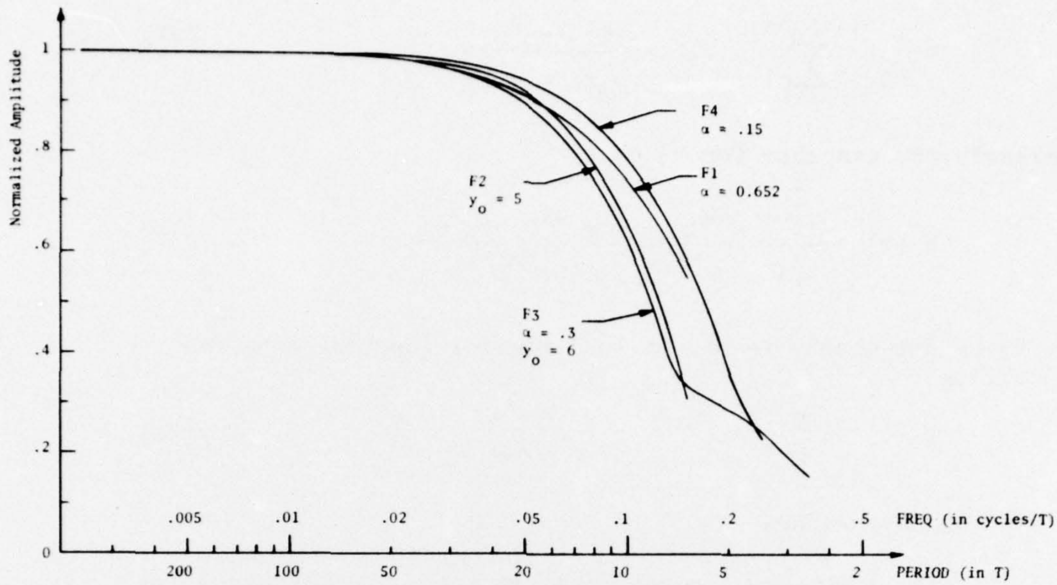

FIGURE 4 - Frequency response functions for response time equal to 5 T

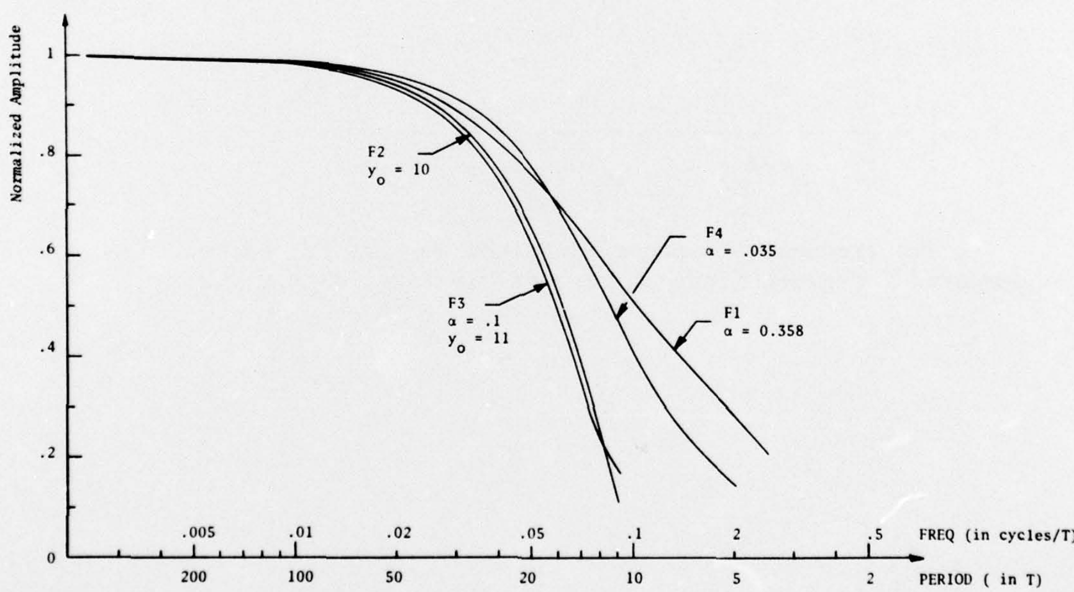

FIGURE 5 - Frequency response functions for response time equal to 10 T

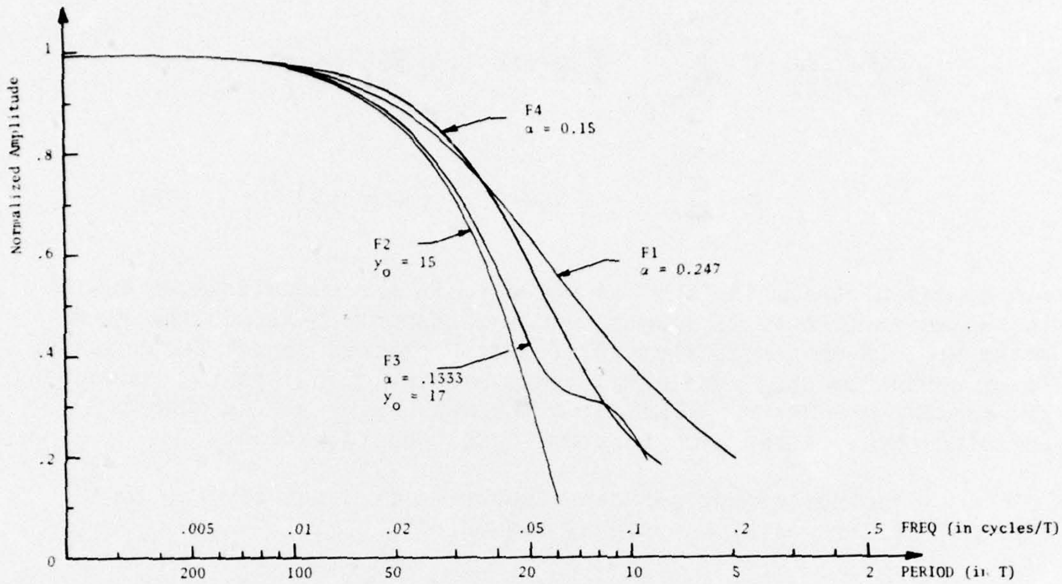

FIGURE 6 - Frequency response functions for response time equal to 15 T

4.0 NUMERICAL SIMULATION OF DIGITAL FILTERS

To study the efficiency of the filters in noise reduction and their performance characteristics in terms of the detection probability and false alarm rate, we have carried out numerical simulations on a general-purpose computer. The details of the simulation are described in a separate document [3]; here, only the main results are described.

4.1 Noise reduction

As mentioned before, one of the main objectives of introducing the DFCU is to extract from a given incoming signal sequence the background signal level for comparison with the current data point. The extent to which a given filter achieves this can be measured by the corresponding noise reduction. For this purpose we may consider the four types of filters with response times equal to 5T, 10T and 15T, and two different input signal sequences. The first sequence is given by:

$$M(y) = C + n(y) \qquad (37)$$

where C is a constant and $n(y)$ is a random gaussian noise with a standard deviation equal to $\sqrt{C}$. To compare the noise in the output sequence with that in the input sequence the ratio:

$$\frac{\text{var } \overline{M}(y)}{\text{var } M(y)} = \sum_{y'=y_0}^{N+y_0} \left(M(y') - \langle \overline{M}(y') \rangle \right)^2$$

$$\div \sum_{y'=y_0}^{N+y_0} \left(M(y') - \langle M(y') \rangle \right)^2 \qquad (38)$$

can be calculated. The initial point y_0 in the summations of Eq. 38 is chosen so that it is larger than the response time and the final point $N+y_0$ is chosen so that the results obtained do not show significant variation when additional terms are included into the summations. The angular bracket $\langle \; \rangle$ indicates the mean value of the quantity enclosed taken in the time interval under consideration.

Another comparison can also be carried out for the four types of filters using two signal sequences:

$$M_1(y) = C \sin \frac{2\pi y}{Y} \qquad (Y = 30)$$

and

$$M_2(y) = C \sin \frac{2\pi y}{Y} + n(y)$$

With the filtered output the mean square error, (MSE) as defined below, is calculated:

$$\text{MSE} = \sum_{y'=y_0}^{N+y_0} \left(\overline{M}_2 (y') - \overline{M}_1 (y') \right)^2$$

$$\div \sum_{y'=y_0}^{N+y_0} M_1^2 (y) \qquad (39)$$

The results of these calculations are given in Tables I to III which show, that for an equal response time, F2 and F3 (for the best combination of the parameters α and Y_0) give about the same noise reduction with a constant or sinusoidal background. Although F4 does not reduce the noise as effectively as F2 or F3, it does give better results than F1. However, from the phase transfer functions of these filters (Tables I to III) it can be seen that, because of their larger phase shifts, F2 and F3 will tend to introduce more errors than F1 when a quickly varying background is present.

Apart from separating the background signal from noise it is desirable that the DFCU should establish and update the detection threshold. Since the background is assumed to be slowly varying we may calculate the root-mean-square noise level σ from the equation

$$\sigma(y) = \sqrt{\frac{1}{y+1} \sum_{y'=y-y_0}^{y} \left[M(y') - \overline{M}(y' - 1) \right]^2} \qquad (40)$$

or in an equivalent but recursive manner:

$$\sigma(y) = \sqrt{\sigma^2 (y - 1) \frac{y}{y+1} + \frac{1}{y+1} \left[M(y) - \overline{M}(y - 1) \right]^2} \qquad (41)$$

The choise of a suitable value of N in Eq. 40 is determined by the consideration, on the one hand, that it should be sufficiently large so that $\sigma(y)$ is not sensitive to the change in the length of the time interval in the summation and, on the other hand, that it should be kept as low as possible to reduce the 'warm-up' time of the whole system.

To update the value of $\sigma(y)$ we may either calculate a separate $\sigma(y)$ from the incoming signal while the $\sigma(y)$ already available is kept until the calculation is completed. A simpler method is to update σ according to the equation:

$$\sigma(y) = \sqrt{\frac{N}{N+1} \sigma^2(y) + \frac{1}{N} \left(M(y) - \overline{M}(y - 1) \right)^2} \qquad (42)$$

Both methods have been tested and they gave essentially the same results.

4.2 Detection probability and false alarm rate

With a given value of $\sigma(y)$, an incoming signal $M(y)$ can be compared to the filtered signal $M(y - 1)$ to determine which of the following two relations is satisfied:

$$|M(y) - \overline{M}(y - 1)| < k\sigma \qquad (43\ a)$$

$$|M(y) - \overline{M}(y - 1)| \geq k\sigma \qquad (43\ b)$$

To assess the performance of the different types of filters we may assume a constant background with a gaussian noise added to a randomly occurring signal of such a magnitude that the signal-to-noise ratio is equal to three. The detection probabilities and the false alarm rates were found for different values of k in

relations 43 a and 6. The results, which are averages over eight trials, are shown in Table IV. For all the filters, the probabilities of detection and the false alarm rates are high for small values of k. Increasing the value of k lowers the false alarm rate and the detection probability at the same time. The differences in the performance of the various types of filters can be compared by calculating the number of the random signals detected minus the number of false alarms in a given period of time (see Table IV). It should be noted that different values of k are required to optimize the different filters.

TABLE I

RELATIVE PERFORMANCE OF FILTERS
(RESPONSE TIME 5 T)

Filter description	Coefficients	Transfer function: A = normalized amplitude, P = phase (degrees), Y = Period					Noise reduction $\frac{\text{VAR } \overline{M}\ (y)}{\text{VAR } M\ (y)}$	Mean square error %
		Y=	20T	30T	40T	50T		
F1	α = 0.652	A=	0.905	0.955	0.973	0.983	0.337	2.047
		P=	-19.57	-12.45	-9.53	-7.69		
F2	y_0 = 5.000	A=	0.904	0.957	0.975	0.984	0.237	1.335
		P=	-36.00	-24.00	-18.00	-14.40		
F3	α = 0.300	A=	0.877	0.951	0.968	0.982	0.245	1.406
	y_0 = 6.000	P=	-29.98	-19.81	-14.93	-11.97		
F3	α = 0.500	A=	0.890	0.955	0.970	0.983	0.289	1.712
	y_0 = 7.000	P=	-22.66	-15.51	-11.75	-9.45		
F3	α = 0.600	A=	0.903	0.957	0.974	0.984	0.320	1.927
	y_0 = 8.000	P=	-19.28	-13.38	-10.17	-8.19		
F4	α = 0.150	A=	0.934	0.973	0.983	0.990	0.294	1.771
		P=	-20.55	-13.84	-10.42	-8.35		

TABLE II

RELATIVE PERFORMANCE OF FILTERS
(RESPONSE TIME 10 T)

Filter description	Coefficients	Transfer function A = normalized amplitude P = phase (degrees) Y = Period					Noise reduction VAR $\overline{M}$ (y) / VAR M (y)	Mean square error %
		Y=	20T	30T	40T	50T		
F1	α = 0.358	A=	0.753	0.866	0.916	0.945	0.207	1.123
		P=	-34.10	-24.71	-19.48	-15.97		
F2	y_0 = 10.000	A=	0.631	0.833	0.898	0.938	0.131	0.587
		P=	-81.46	-54.00	-40.50	-32.40		
F3	α = 0.100	A=	0.601	0.808	0.888	0.928	0.129	0.570
	y_0 = 11.000	P=	-69.07	-47.14	-35.74	-28.72		
F3	α = 0.250	A=	0.677	0.845	0.906	0.940	0.168	0.841
	y_0 = 12.000	P=	-44.89	-32.59	-25.16	-20.38		
F3	α = 0.300	A=	0.707	0.851	0.906	0.942	0.183	0.955
	y_0 = 14.000	P=	-38.65	-29.05	-22.67	-18.47		
F3	α = 0.350	A=	0.747	0.863	0.915	0.945	0.203	1.100
	y_0 = 19.000	P=	-34.85	-25.31	-19.92	-16.31		
F4	α = 0.035	A=	0.758	0.892	0.933	0.959	0.176	0.904
		P=	-45.51	-31.42	-23.93	-19.28		

TABLE III

RELATIVE PERFORMANCE OF FILTERS
(RESPONSE TIME 15 T)

Filter description	Coefficients	Transfer function: A = normalized amplitude, P = phase (degrees), Y = Period					Noise reduction $\frac{VAR\ \overline{M}\ (y)}{VAR\ M\ (y)}$	Mean square error %
		Y=	20T	30T	40T	50T		
F1	α = 0.247	A=	0.615	0.766	0.844	0.893	0.153	0.742
		P=	-43.52	-34.55	-28.16	-23.53		
F2	y_0 = 15.000	A=	0.301	0.638	0.785	0.859	0.093	0.305
		P=	-126.00	-84.00	-63.00	-50.40		
F3	α = 0.003	A=	0.254	0.604	0.761	0.844	0.091	0.292
	y_0 = 16.000	P=	-109.29	-79.42	-60.38	-48.55		
F3	α = 0.133	A=	0.408	0.661	0.790	0.861	0.111	0.437
	y_0 = 17.000	P=	-63.80	-53.86	-42.80	-35.04		
F3	α = 0.167	A=	0.475	0.684	0.802	0.870	0.121	0.512
	y_0 = 18.000	P=	-55.07	-47.26	-38.09	-31.39		
F3	α = 0.200	A=	0.538	0.722	0.822	0.882	0.134	0.615
	y_0 = 19.000	P=	-50.22	-41.46	-33.60	-27.80		
F3	α = 0.233	A=	0.594	0.748	0.833	0.889	0.147	0.699
	y_0 = 24.000	P=	-44.97	-36.35	-29.75	-24.82		
F4	α = 0.015	A=	0.562	0.766	0.858	0.907	0.125	0.539
		P=	-65.00	-47.71	-37.05	-30.10		

TABLE IV

RELATIVE PERFORMANCE OF FILTERS

(Average over 8 trials, each with 50 events to be detected)

k		F1	F2	F3	F3	F4
		$\alpha = 0.358$	$y_0 = 10.000$	$\alpha = 0.100$ $y_0 = 11.000$	$\alpha = 0.530$ $y_0 = 19.000$	$\alpha = 0.035$
1.8	D	50.0	50.0	50.0	50.0	50.0
	F	35.8	33.8	27.8	38.4	34.0
	D - F	14.2	16.2	22.2	11.6	16.0
2.0	D	50.0	50.0	50.0	50.0	50.0
	F	23.4	24.2	24.2	23.0	29.4
	D - F	26.6	25.8	25.8	27.0	29.6
2.2	D	50.0	50.0	50.0	50.0	50.0
	F	19.8	16.4	15.6	18.8	20.4
	D - F	30.2	33.6	34.4	31.2	29.6
2.4	D	50.0	50.0	50.0	50.0	50.0
	F	6.0	7.6	8.0	11.4	9.0
	D - F	44.0	42.4	42.0	38.6	41.0
2.6	D	41.0	50.0	49.4	41.6	47.6
	F	7.2	4.8	5.0	5.4	6.8
	D - F	33.8	45.2	44.4	36.2	40.8
2.8	D	38.4	48.8	35.6	39.2	29.6
	F	4.0	2.6	3.2	5.4	3.0
	D - F	34.4	46.2	32.4	33.8	26.6

D: Detected

F: False alarm

D-F: Detected minus False alarm

5.0 CONCLUSIONS

From the noise reduction simulation and the calculation of the detection probability and the false alarm rate discussed above, we can see that among the four types of filters F2 gives the best performance. The results obtained for F3 are quite close to those obtained for F2, and those for F4 show that it is the least effective. For the purposes of building a testing model it seems only necessary to consider F1 and F2. The great advantage of F1, in spite of its poorer performance, is the simplicity of its implementation. Its characteristics are completely determined by a parameter K, which can easily be varied. For F2, on the other hand, the only variable parameter is the memory capacity once the number of sampling points per scan is fixed. This is a significant drawback in comparison with F1.

The results presented in Sec. 4.0 for the detection probability and false alarm rate also show that F2 gives a better performance than F1. To reduce the false alarm rate without decreasing at the same time the detection probability, we may decide to trigger an alarm only after observing an anomalous variation in the incoming signals over a number of consecutive scans in a correlated space sector. To do this, only minor modifications are required in the signal processing.

In this study we have not considered the question of making use of more specific information on targets and backgrounds. This is one area in which further work would help to enhance the discriminating power of the intrusion alarm.

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